DESIGN OF A VHF DIRECTION-FINDER FOR TORNADO DETECTION

By

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CHAPTER I

INTRODUCTION

It has been shown that tornadic clouds generate pulses of electromagnetic energy. Most of the investigation carried on at the Oklahoma A. and M. College under the supervision of Dr. H. L. Jones has been at video frequencies. The amount of energy in the VHF spectrum is still unknown and considerable investigation will be necessary to determine this.

The main concern of this paper is the design of direction-finding equipment that may be used in the VHF band, that is, the 144 mc. amateur band. At these frequencies the man-made static has decreased to such an extent that the noise in the receiver is a determining factor for the minimum signal that can be received. The noise figure is treated quite thoroughly in Chapter III because it is the biggest problem in the obtaining of the maximum available signal-to-noise ratio. The receiver is a BC 624A converted by Mr. Lin Seow to be tunable over the 144-148 mc. band. In order to increase the overall gain and improve the signal-to-noise ratio a Wallman cascode amplifier is added to the receiver. In this receiver the automatic-gain-control voltage is used as an indicator. If the signal is too low an oscilloscope is placed across the second-detector load resistance and, with suitable amplification, the signal can be viewed. This setup is designed to aid in determining approximately the amplitude and type of signals to be expected.

The antenna is designed to give a fairly wide bandwidth so that if the receiver had to be designed for a wider band-pass, the antenna would still be usable. There is quite a bit more experimentation to be done on the antenna since at frequencies above two meters it is impossible to measure the current in each element, let alone the phase. Some suggestions for further tests have been included in the conclusions. Material is provided for the design of the direction-finder with suggestions for improving its performance. Since it is impossible to list all references, especially in antenna theory, a number of books were cited in which the reader may obtain further references to original studies.

CHAPTER II

ANTENNAS

Introduction

An antenna is a "transformer" that converts electromagnetic waves into guided waves or vice versa. The directional patterns are the same for transmitting and receiving, but the optimum pattern for transmitting may not be optimum for receiving. For transmission, the optimum pattern is usually one that puts the maximum energy in the required direction. For reception the optimum pattern is one that gives the maximum signal-to-noise ratio. The patterns may be the same but not necessarily.

The noise generated in the antenna by the radiation resistance is fixed, but the external noise depends on its direction of arrival. For direction-finders there should be a minimum of side lobes to prevent spurious readings and the input coupling should be adjusted for maximum signal-to-noise ratio.[1]

Binomial Array

The antenna that meets the requirement of practically no side lobes is the binomial array. It can be made unidirectional and appears to be the most useful. Due to the large background required for a through understanding of antenna theory, it was decided only to include the actual design and a few pertinate facts about it, and give references at the end of the thesis for the use of the reader who wishes further information.

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[1] This is best done by a noise generator unless the antenna noise is much greater than the receiver noise.

The binomial array was first brought to the attention of the author in a QST article by Warren M. Andrews.[2] The article gave dimensions for a three-element end-fire array on 14 mc. It was decided at first to just scale down the antenna and use it for 144 mc., but theoretically this is incorrect.

The essentials of the array are as shown in Figure 1.

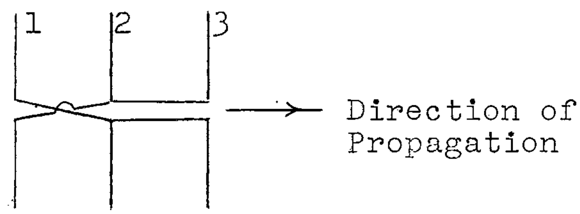

Figure 1

The current ratios are

$$2I_1 = I_2 = 2I_3$$

The elements are half-wave length and are spaced a quarter-wave length apart. The current in element 1 leads that of the center element by 90°, and the current in element 3 lags that of the center element by 90°, since the array is fed at the center element. Elements 1 and 3 are three-element folded-dipoles, while the center is a two-element dipole.

<u>Theory of the Binomial Array</u>

Research on the theory of antennas disclosed the interesting fact that the mutual impedance between the elements cause a considerable change in the input impedance of the elements.

In a recent article by Ronold King[3] it is stated that if

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[2] Reference 1

[3] Reference 13

the center-driven antenna length is near a half-wave length, then the array may be analyzed in terms of the self-impedance of isolated antennas and mutual impedances of two isolated coupled antennas and that the error will be negligible. Using this as the basis for computations the input impedances of the different elements may be found.

The voltage at the terminals of element 1 is

$$V_{10} = I_1 Z_{11} + I_2 Z_{12} + I_3 Z_{13}$$

since

$$\frac{I_2}{I_1} = \frac{2}{j} = -j2$$

$$\frac{I_3}{I_1} = \frac{-j}{j} = -1$$

then

$$Z_{10} = Z_{11} - j2Z_{12} - Z_{13}$$

From Reference 15, page 266, the value for the self- and mutual impedances can be found.

Since the antenna is cut to a half-wave physical length there results,

$$Z_{11} = 70 + j0 \qquad Z_{21} = 38 - j38 \qquad Z_{31} = -13 - j26$$

$$Z_{12} = 38 - j38 \qquad Z_{22} = 70 \qquad Z_{32} = 38 - j38$$

$$Z_{13} = -13 - j26 \qquad Z_{23} = 38 - j38 \qquad Z_{33} = 70$$

therefore

$$Z_{10} = 7 - j50 \text{ ohms.}$$

Element 2 can be represented as,

$$V_{20} = I_1 Z_{21} + I_2 Z_{22} + I_3 Z_{32}$$

$$Z_{20} = \frac{jZ_{21}}{2} + Z_{22} - \frac{j}{2} Z_{23}$$

$$Z_{20} = 70 \text{ ohms.}$$

For element 3 there results,

$$V_{30} = I_1Z_{31} + I_2Z_{32} + I_3Z_{33}$$

$$Z_{30} = -Z_{31} + j2Z_{32} + Z_{33}$$

$$Z_{30} = 159 + j10_2$$

Therefore the input impedance in elements 1 and 3 is very much different from that of an isolated dipole but remains the same for element 2.

$$Z_{10} = 7 - j50 \text{ ohms}$$

$$Z_{20} = 70 \text{ ohms}$$

$$Z_{30} = 159 + j102 \text{ ohms}$$

These are the impedances that will appear at the input to the elements if the current distributions are as assumed. For correct loading the phasing lines must be used for impedance matching.

Impedance of Phasing Lines

The input impedance of element 1 as a three-element folded dipole is

$$Z_1 = 9Z_{10} = 63 \text{ ohms.}$$

The input impedance of element 2 as a two-element folded dipole is

$$Z_2 = 4Z_{20} = 280 \text{ ohms.}$$

The input impedance of element 3 as a two-element folded dipole is

$$Z_3 = 4Z_{30} = 636 \text{ ohms.}$$

The power consumed in the elements are

$$P_1 = I_1^2R_1 = 4I_2^2R_1 = I_2^2(252)$$

$$P_2 = I_2^2(300)$$

$$P_3 = I_2^2(2540)$$

Since the antenna is fed at the input of the center element, the phasing elements must transform the impedances so that they will give the correct currents from the same voltage source, and there results

$$P_1 = \frac{E_2^2}{R_{11}} = I_2^2(252)$$

$$P_2 = \frac{E_2^2}{R_{22}} = \frac{E_2^2}{280} = I_2^2(280)$$

$$P_3 = \frac{E_2^2}{R_{33}} = I_2^2(2540)$$

where R_{11}, R_{22}, and R_{33} are the transformed impedances.

Solving these equations for the transformed impedances it follows that

$$R_{11} = 314 \text{ ohms}$$

$$R_{33} = 30.9 \text{ ohms.}$$

The characteristic impedances of the lines will then be

$$Z_{12} = (63)(314) = 141 \text{ ohms}$$

$$Z_{23} = (636)(30.9) = 140 \text{ ohms.}$$

The characteristic impedance of the line in terms of its physical dimensions is

$$Z_o = 120 \cosh^{-1} \frac{D}{d}$$

where D is the distance between the centers and

d is the diameter of the conductor.

Solving for $\frac{D}{d}$ there results

$$\frac{D}{d} = \cosh \frac{Z_o}{120} = 1.89$$

If #8 wire is used d is .1285 inches in diameter and the conductors should be spaced about one-fourth inch apart. Twin-lead can not be used for the transmission lines between the elements because the velocity of propagation in the line is not the same as in free space and would cause incorrect phasing. The open wire line does not quite do this, but it is so close that the out-of-phase effect is negligible.

Folded Dipole[4]

Since the folded dipole is used in the antenna, a short discussion of this subject is in order. Any folded dipole has a higher impedance at the input terminals than a simple dipole employed at the same place in any antenna or antenna array. This property of impedance transformation has led to the increased use of folded dipoles, especially at very high frequencies.

The folded dipole may be considered as two series-connected stub transmission lines, a quarter-wave in length, shorted at one end. This is the transmission line mode of the antenna, which radiates only a negligible amount of energy, and aids in increasing the bandwidth. At the resonant frequency the dipole resistance is in parallel with the input impedance of the transmission line, which is usually a resistance of very high value. Below resonance the antenna impedance becomes capacitive, but the transmission line impedance becomes inductive and the parallel combination tends to remain nearer unity power factor than does the antenna alone. Conversely, above resonance the antenna becomes inductive and the line impedance becomes capacitive so

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[4] Reference 4; Reference 12, pp 534; Reference 10; Reference 14, pp. 224; Reference 15, pp. 415; Reference 25.

that compensation is again obtained. At single points above and below the resonant frequency perfect susceptance compensation is obtained so that there are three points of pure resistance input. It follows that the input resistance at the stub-matched or anti-resonant points will be considerably higher than at the resonant frequency.

At resonance the effect of the transmission line mode may be neglected and the antenna or in-phase currents only need be considered. Assuming that the radiation from a folded dipole at the same place, it is possible to compute the transformation ratio, and consequently the impedance, at the feeding point if the ratio of currents in the elements of the folded dipole is known. The current at the feed point is designated as I_1, and the current at the center of the auxiliary element as I_2. In any array in which the fed element is a simple dipole, let the input resistance at the feed point be R_o. When the simple dipole is replaced by a folded dipole, let the new resistance be R_1. Then, with the assumption of equal radiation there results

$$I_1^2 R_1 = (I_1 + I_2)^2 R_o.$$

The folded dipole, therefore, gives the resistance transformation ratio, u, as

$$u = \frac{R_1}{R_o} = \left(\frac{I_2}{I_1} + 1\right)^2 = (n + 1)^2$$

where n is the current ratio $\frac{I_2}{I_1}$.

For a two-element folded-dipole the current ratio is[5]

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[5] Reference 10.

$$n = \frac{\text{Log } \frac{s}{a_1}}{\text{Log } \frac{s}{a_2}}$$

where s is the distance between the centers of the elements and a is the radius of the elements.

The impedance ratio is

$$u = \left(\frac{\text{Log } \frac{s^2}{a_1 a_2}}{\text{Log } \frac{s}{a_2}} \right)^2$$

provided

$$\frac{a_2}{a_1} \geqq 1 \text{ and } \frac{s}{a_2} \geqq 2.5$$

$$\frac{a_2}{a_1} < 1 \text{ and } \frac{s}{a_1} \geqq 2.5$$

Conclusions

To most critical readers this treatment of antennas is inadequate, but to cover the subject thoroughly it would require a prohibitive amount of space. The books by J. D. Kraus[6] and E. C. Jordan[7] cover the subject of antennas as well as any text available. E. B. Moullin,[8] in his book, gives an excellent slant on the British point of view. Antenna measurements procedure is covered in the books of J. D. Kraus and E. B. Moullin.

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6 Reference 21.

7 Reference 12.

8 Reference 17.

CHAPTER III

NOISE FIGURES

Introduction

To be useful, any receiver must be connected to a source of signals such as an antenna. Even if a receiver could be produced that had no internal sources of noise, noise would still be induced into the system from the antenna, and weak signals would have to compete with this noise. If the receiver does introduce additional noise, the signal must be correspondingly stronger. These concepts may be made quantitative by the use of the noise figure.

The noise figure[1] is not a measure of the excellence of the output signal-to-noise ratio, but merely a measure of the degradation suffered by the signal-to-noise ratio as the signal and noise pass through the network in question. If the input signal and noise are extremely large compared to the amplifier noise, relatively little degradation is suffered, and the noise figure is good.

Friis[2] defines the noise figure, F, of a network as the ratio of the available signal-to-noise ratio at the signal generator terminals to the available signal-to-noise ratio at its output terminals. To understand the meaning of this definition a few basic concepts will be defined.

Available Power

The most commonly used representation of an arbitrary electrical generator is a constant-voltage source in series with a

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[1] Reference 8, p. 1208.

[2] Reference 7.

constant internal impedance. It is possible, by Thevenin's Theorem,[3] to represent any linear network as shown in Figure 2.

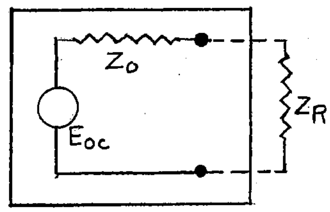

Figure 2.

The generated voltage E_{oc} is the open circuit voltage at the output terminals and Z_o is the impedance of the network looking back from the terminals, with all generators replaced by impedances equal to their internal impedance. If E_{oc} and Z_o are known, along with one of the following four quantities concerning the load, the other three may be computed: the impedance of the external load, the voltage across it, the current through it, and the power dissipated in it.

Friis[4] defines the "available power" as the maximum power that the generator can deliver to the load. It is independent of the load, by definition, and depends only on the characteristics of the generator. The maximum power is obtained only when the load is matched to the generator, that is, when the load impedance, Z_R, and the internal generator impedance Z_o are conjugates.[5] The maximum power absorbed from the generator, that is the "available power", is

$$P_a = \frac{E_{oc}^2}{4R_o}$$

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3 Reference 5, p. 47.

4 Reference 7.

5 Reference 5, p. 49-50.

where R_o is the resistive component of the internal generator impedance Z_o.

Another method of representing an electrical generator is, as suggested by Norton's Theorem,[6] by an infinite-impedance constant-current generator, I_{sc}, shunted by the internal impedance, Z_o. The generated current, I_{sc}, is the short-circuit current that would flow if the output terminals were short-circuited. The "available power" is

$$P_a = \frac{I_{sc}^2 R_o}{4} = \frac{I_{sc}^2}{4G_o}$$

where G_o is the conductive component of the internal generator admittance, Y_o. By definition, $Y_o = \frac{1}{Z_o}$.

Available Noise Power

Using the definition of "available power", the "available noise power" may be defined as the maximum power delivered to the load from the noise generator. As an example, the thermal noise of a resistor as a noise generator will be computed. It is well known that a resistor, R, at an absolute temperature, T, produces a noise voltage across its terminals. It is usually represented as a generator having an internal noise-free resistance and an open-circuit mean-squared voltage of

$$E^2 = 4kTBR$$

where k is Bolzmann's constant equal to 1.380×10^{-23} joules per degree Kelvin,

B is the small bandwidth over which E is measured in

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6 Reference 5, p. 48.

cycles per second,

T is the room temperature in degrees Kelvin, and

R is the resistance in ohms.

The bandwidth appears in the equation because noise is not a single frequency phenomenon. The bandwidth is defined as

$$B = \frac{G_p dF}{dG_{po}}$$

where G_p is the actual power gain delivered to the indicating device divided by the available power of the signal source and

G_{po} is the maximum value of G_p. This definition is equivalent to replacing the power-frequency curve by a rectangular curve having the same area under it and a maximum value of G_{po}. For all practical purposes, B is very nearly equal to the bandwidth between the half power points, except the pass-band of a single tuned circuit which is $\frac{\pi}{2}$ or 1.57 times the half-power bandwidth. The available noise power is then

$$N = \frac{E_{oc}^2}{4R_o} = \frac{4kTBR_o}{4R_o} = kTB.$$

This is the noise power delivered by the resistor to a noise-free resistor of equal resistance. It is not necessarily the actual noise power dissipated in either of two resistors of equal resistance connected in parallel.

Available Gain

The "available gain" is defined as the ratio of "available output power" from the network to the "available signal power" from the generator. Let it be assumed that the generator is connected to a network and the network is connected to a load.

The generator will deliver power to the network and the network will deliver power to the load. With the load disconnected, the output terminals of the network displays an internal impedance. If the external load matches this impedance, the power delivered to the external load is the "available output power" of the network. The "available output power", therefore, will be independent of the load, but not independent of the way in which the signal generator is coupled to the network.

The available gain of a cascade of networks may now be computed. Given a cascade of networks fed by a signal generator, the available gain, G_1, of the first network is measured by opening the cascade between networks 1 and 2. The available gain of network 2 can be measured by opening the cascade between network 1 and 2, and 2 and 3, and by employing a signal generator of internal impedance equal to the impedance seen looking into the output of network 1 when the original signal generator is still connected. This gain can be called G_2. This process should be carried out for each network in turn. The total gain is the product of the "available gains", and there results

$$G_1G_2G_3\ldots G_n = \left(\frac{S_1}{S_g}\right)\left(\frac{S_2}{S_1}\right)\left(\frac{S_3}{S_2}\right)\cdots\cdot\left(\frac{S_n}{S_{n-1}}\right)$$

where S_n is the available output power of the nth network, when measured as above, and S_g is the available generator power. Simplifying there results,

$$G_1G_2G_3\ldots G_n = \frac{S_n}{S_g}$$

This result is valid only when the "available gain" of each element is measured with a signal generator whose internal impedance is the impedance of the entire network preceding the element under measurement.

Calculation of Available Gain

For future use the "available gain" of a signal generator with a resistor in series should be computed. The "available power output" may be computed by the use of Thevenin's Theorem as indicated in Figure 3.

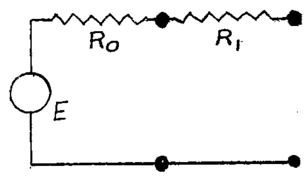

Figure 3.

If the open-circuit output voltage is taken as E, and the output impedance is taken as $R_1 + R_o$, the maximum power output then will be

$$S_n = \frac{E_{oc}^2}{4R_o} = \frac{E^2}{4(R_1+R_o)}$$

The "available signal power" from the generator alone is

$$S_g = \frac{E^2}{4R_o}$$

and by definition the "available gain" is

$$G = \frac{S_n}{S_g} = \frac{\dfrac{E^2}{4(R_1+R_o)}}{\dfrac{E^2}{4R_o}} = \frac{R_o}{R_1+R_o}$$

In a similar way the "available gain" of a signal generator with a resistor in parallel may be computed from the circuit shown in Figure 4.

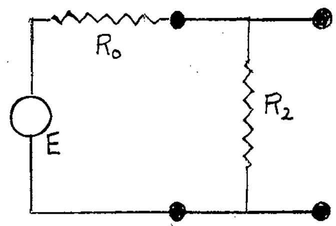

Figure 4.

It is evident that the open-circuit voltage will be

$$E_{oc} = \frac{ER_2}{R_o+R_2}$$

and the output impedance will be

$$Z_{out} = \frac{R_oR_2}{R_o+R_2}$$

and hence the maximum power output becomes

$$S_n = \frac{E_{oc}^2}{4R_{out}} = \frac{E^2\left(\frac{R_2}{R_o+R_2}\right)^2}{4\left(\frac{R_o\ R_2}{R_o+R_2}\right)} = \frac{E^2\left(\frac{R_2}{R_o+R_2}\right)}{4R_o}$$

The "available signal power" from the generator will be

$$S_g = \frac{E^2}{4R_o}$$

and hence the "available gain" becomes

$$G = \frac{S_n}{S_g} = \frac{\dfrac{E^2\left(\frac{R_2}{R_o+R_2}\right)}{4R_o}}{\dfrac{E^2}{4R_o}} = \frac{R_2}{R_o+R_2}$$

The ideal transformer as shown in Figure 5, will be considered next. Since it has no dissipation, there should be no change in the "available gain".

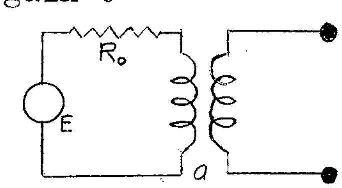

Figure 5

If the transformation ratio is a, the output open-circuit voltage will be aE, and the output impedance becomes a^2R_o. The "available power output" then will be,

$$S_n = \frac{E_{oc}^2}{4R_{out}} = \frac{a^2E^2}{4a^2R_o} = \frac{E^2}{4R_o} ,$$

while the "available signal output" is

$$S_g = \frac{E^2}{4R_o}$$

and hence the "available gain" is

$$G = \frac{S_n}{S_g} = \frac{\frac{E^2}{4R_o}}{\frac{E^2}{4R_o}} = 1 ,$$

From this result it is evident that the "available gain" for a nondissipative, passive network is always unity.

Now if the concept of "availiable gain" is expanded to include active network, it will be necessary to consider an amplifier driven by a generator with a resistor, R_3, across its input terminals, as shown in Figure 6.

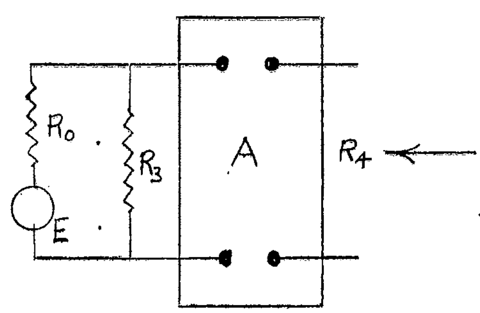

Figure 6.

The amplifier has an infinite input impedance and an open-circuit voltage gain of A. The output impedance, R_4, is assumed to be independent of the impedances connected to the input.

The "available output power" becomes,

$$S_n = \frac{E_{oc}^2}{4R_{out}} = \left(\frac{\frac{AER_a}{R_3+R_a}}{4R_4}\right)^2 ,$$

and the "available signal power" will then be,

$$S_g = \frac{S_n}{S_g} = \frac{\frac{\left(\frac{AER_3}{R_3+R_o}\right)^2}{4R_4}}{\frac{E^2}{4R_4}} = \frac{R_oA^2}{R_4} \quad \frac{R_3}{R_3+R_4}$$

The foregoing derivations of "available gain" can be used in conjuction with the cascade-network theorem in the previous section to find the "available gain" of any network. Serious errors will result if the proper procedure is not used in computing the "available gain" of an element in the cascaded network.

Noise Figures

Sufficient background has now been presented to discuss intelligently the concept of noise figures. The first case to be considered will be that of an ideal amplifier, that is, an amplifier which is noise free and has effectively an infinite input impedance. The gain of the ideal amplifier is made sufficiently large so that the amplified noise from the signal-generator source is very much larger than the noise generated in any resistive load connected to the output terminals. A bandwidth, B, is assumed and all impedances involved will be real at the frequencies within the bandwidth. This assembly is shown in Figure 7.

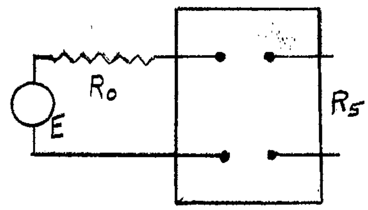

Figure 7.

From the previous material, it is evident that the "available signal power" will be

$$S_g = \frac{E^2}{4R_o} ,$$

and that "available output power" due to the signal is

$$S_n = \frac{E^2A}{4R_5} .$$

The "available gain", then becomes,

$$G = \frac{S_n}{S_g} = \frac{\frac{E^2A}{4R_5}}{\frac{A^2}{4R_o}} = \frac{A^2R_o}{R_5}$$

It is customary to consider the noise generator, connected to the input of an amplifier, as a resistor equal to the resistance of the source, for example, the radiation resistance of an antenna. The "available output-noise power" can be obtained by extending the previous analysis. If the source voltage is

$$E^2 = 4kTBR_o ,$$

the "available noise output power" becomes

$$N = \frac{4kTBR_oA^2}{4R_5} = kTB\left(\frac{R_oA^2}{R_5}\right)$$

But since $G = \frac{A^2R_o}{R_5}$, it follows that $N = GkTB$

By definition the "available signal-output power" is,

$$S_n = GS_g,$$

and by comparing the last two equations it may be concluded that the "available noise power" from the generator is kTB, which was previously demonstrated by another method.

From the definition of the noise figure as stated in the first section, it may be written symbolically as

$$F = \frac{\frac{S_g}{kTB}}{\frac{S_n}{N}} = \frac{S_g}{kTB}\ \frac{N}{S_n} = \frac{N}{GkTB}$$

This equation affords a method of computing noise figures. Solving for N, the expression for the "available noise-output power", in terms of the noise figure, gain, and available noise-input power there results

$$N = FGkTB.$$

The question that immediately arises is the relative contributions of the signal generator and the noise from the network to the "available noise output". Since the "available noise power" from the generator is kTB, the "available noise output power" from the generator is GkTB. It follows, therefore, that the contribution from the network itself is (F-1)GkTB.

This separation of the "available noise-output power" into two parts provides a method by which an expression for the noise figure of a cascade network may be written in terms of the noise figure of the individual networks. This becomes,

$$F_{1n} = F_1 + \frac{(F_2-1)}{G_1} + \frac{(F_3-1)}{G_1G_2} + \ldots \frac{(F_n-1)}{G_1G_2\ldots G_n} \quad ,$$

where G_n denotes the "available gain" of the n^{th} network. Usually in a receiver employing several stages it is seldom necessary to consider the noise figure of more than the first two stages unless the "available gain" of the first stage or stages is a loss. The equation may then be rewritten as

$$F_{1n} = F_1 + \frac{(F_a-1)}{G_1},$$

where F_a is the noise figure of the rest of the stages.

In this formula it was assumed that the bandwidths of each of the networks, were the same. If this is not the case, there results

$$F_{in} = F_1 + \frac{(F_a-1)B_a}{G_1 B_1}$$

where the subscripts have the same meaning as before.

<u>Sources of Noise in Tubes</u>[7]

The principal sources of noise in tubes are

1. Shot effect (temperature-limited emission).
2. Reduced shot effect.
3. Flicker effect.
4. Collision ionization.
5. Random division of current between electrodes.
6. Inducted noise at ultra-high frequencies.
7. Faulty tube construction.

1. Shot Effect. Shot effect is the noise associated with random emission in a tube when the emission is temperature-limited. Since most tubes are not operated so that their emission is not

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7 Reference 22, p. 305.

temperature limited, this noise is not serious even though it probably is the loudest.

2. Reduced Shot Effect. Most tubes are operated under space-charge-limited conditions which has a very definite "smoothing" action upon shot effect, which is called "reduced shot effect". The smoothing action may be a result of the formation of a virtual cathode in front of the emitting surface which has a potential lower than that of the emitter by a value determined by the mean velocity of emission. Electrons with all velocities are storming this potential hill, and those with velocities greater than the mean velocity will <u>on the average</u> get past the virtual cathode and go on to the plate. Occasionally there will come a group of electrons with a velocity slightly in excess of that needed to get past the virtual cathode. When this occurs, the potential minimum at the virtual cathode is momentarily depressed by the additional space charge and as a result a few electrons that normally would have gotten past the potential minimum fail to do so and are returned to the emitter. This means that for every burst of electrons which might give rise to noise there is a compensating current set up in the opposite direction which tends to cancel the noise produced by the burst. The net result is an over-all reduction in noise that is considerable. The resulting noise power is of the order of 10 percent of that encountered for the same current when the emission is temperature-limited.

3. Flicker Effect. Flicker effect is observed in oxide cathodes. This effect is due to the variations in the activity of the emit-

ting surface and is much noisier than the true shot effect for temperature-limited emission. It's magnitude is also greatly reduced when the emission is space-charge-limited.

4. Collision Ionization. Noise in tubes due to gas molecules being ionized is small unless the positive ion gas current is more than a few hundredths of a microampere. When gas molecules are ionized by collision with emitted electrons, the positive ions formed subsequently liberate small bursts of electrons as they penetrate the virtual cathode in front of the emitting surface. Gas noise appears mostly below 10 mc.

5. Random Division of Current. Random division of current between electrodes contributes to the noise of multielement tubes and makes pentodes three to five times as noisy as the same tube when connected as a triode.

6. Induced Noise at Ultra-High Frequencies. The ultra-high-frequency components of the random fluctuations of the space charge in a tube will induce voltages in the grid circuit, which in turn will react back upon the space-charge flow. This effect is only important for frequency components, above 30 mc.

7. Faulty Tube Construction. Noise due to faulty tube constrution is always present to a degree. If the filament is inductive a hum will result. Poor insulation, mechanical vibration and dirt on the glass inside of a tube are some of the other noise components that may be minimized by careful construction of tubes.

Equivalent Circuit of an Amplifier

The equivalent circuit of an amplifier may be represented as shown in Figure 8,

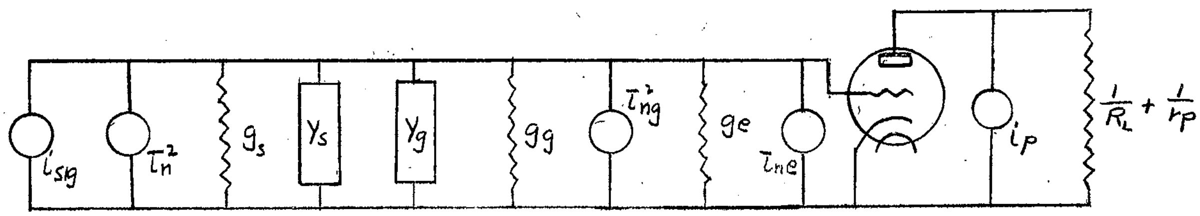

Figure 8.

where

g_s is the transformed source conductance,

Y_s is the transformed source susceptance,

g_g is the conductance due to passive tube loading and circuit losses,

Y_g is the susceptance due to interelectrode and circuit capacitance,

r_p is the tube plate resistance,

R_L is the load resistance, and

g_e is the conductance due to electronic or transit-time loading.

To simplify the analysis it will be assumed that the effects of the various tube and circuit capacitances are eliminated by resonance methods. The noise sources associated with R_L are considered as part of the second stage, and, although the tube shown is a triode, it can also be a tetrode or pentode, provided all electrodes other than the control grid and plate are by-passed to ground over the frequency range to be amplified. It is common practice in noise-factor analysis to replace the actual tube with plate-circuit noise current, i_p, by a fictitious noiseless tube having in series with its grid lead a noise voltage

$$E_{eq} = \frac{i_p}{g_m}$$

which, when amplified by the tube, produces in the plate circuit the noise current, i_p. It is then possible to define a purely fictitious resistance, R_{eq}, according to the relation

$$R_{eq} = \frac{E^2_{eq}}{4kTB}$$

where R_{eq} is the equivalent noise resistance. The equivalent tube noise resistance is, therefore, defined as that resistance which, if placed between grid and cathode of a noise-free tube, would produce in the plate current a noise current equal to that which the actual "noisy tube" produces with its grid shorted to ground. It is also assumed that the cathode lead inductance can be considered to be zero or tuned out by a suitable series capacitor.

The following equations may be used in order to determine the approximate values of R_{eq}:

for triode amplifiers,

$$R_{eq} = 2.5/g_m$$

and for pentode amplifiers,

$$R_{eq} = \frac{I_b}{I_b+I_{c2}} \quad \frac{2.5}{g_m} + \frac{20I_{c2}}{g_m^2} .$$

The statistically-independent constant-current thermal-noise generators may be defined in terms of a mean-squared thermal-agitation-noise current. When the transformed source noise current is

$$\overline{i^2_{ns}} = 2kT_aBg_o,$$

then the noise current due to passive tube loading and circuit losses becomes

$$\overline{i^2_{ng}} = 2kTBg_g,$$

and the current due to the electronic or transit-time loading is

$$\overline{i^2_{ne}} = 4kTBg_e.$$

For tubes with oxide-coated cathodes and space-charge-limited emission the ratio of $\frac{T_b}{T_a}$ is approximately 5.[8,9] Further simplification may be had by combining the tube loading due to cathode-lead inductance feedback, the loading due to losses in the input circuit, and the transit-time loading all into a shunt conductance $\frac{1}{R_g}$. The lead inductance can be considered to be zero or tuned out by a suitable series capacitor, for it has been shown[10] that if this were not the case the resulting input loading would have only second-order effects on the amplifier noise factor, since it degenerates tube noise as well as input noise. It has been found that cathode-lead-inductance loading does affect the band-width.

Since even the summary of the properties of the three types of tride input circuits, grounded cathode, grounded grid, and grounded plate, will be quite large, reference is here made to Reference 9, chapter 13 or Reference 10, chapter 4. Some approximate equivalent noise resistances are given in Table I. These are based on the normal values of the transconductance and

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[8] Reference 3.

[9] Reference 18.

[10] Reference 8.

plate and screen currents given in the RCA Tube Handbook. Average values of input and output capacitances, which are valid for the case when the tube has its cathode grounded, are given; these values do not include socket or wiring capacitance. Although the agreement on the value of equivalent noise resistance between theory and experiment is quite good for some types of tubes, such as the 6AK5, it is much less satisfactory for others. The reasons for this discrepency is not yet fully understood but it is believed that the gold plated grid of the 6AK5, with its low emission, may be responsible for its lower noise. In Table II a comparison of the various single-triode input circuits is given.

Double-Triode Input Circuits[11]

The ideal triode input circuit should have a number of necessary characteristics.

1. All the improvement in input stage noise figure over the pentode that is theoretically possible.

2. The contribution of second and later stages to the noise figure no greater than with pentode input stages of the same band width.

3. A circuit that is stable and no more critical in adjustment than with pentode stages of the same bandwidth.

Of the nine possible combinations with two triode tubes only one meets the above requirements. This is the combination of a grounded cathode followed by a grounded grid.

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11 Reference 24, p. 656.

Grounded-Cathode-Triode-Grounded-Grid-Triode Input Configuration

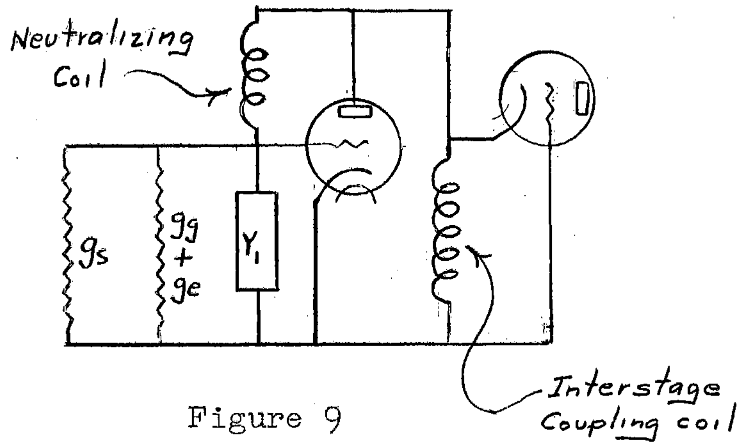

Figure 9

When a grounded-grid stage succeeds the grounded-cathode stage a very large conductance is presented to the plate of the first triode so that it is quite stable. The bandwidth of the interstage coupling will be very large so that there is no need to add additional damping even in the most extreme cases. The output conductance of the first tube is of the same order of magnitude as the optimum source conductance for the second tube, so that the full available power gain of the grounded-cathode triode is utilized.

The exact expression for the noise figure of first two stages combined is[12]

$$F = 1 + \frac{g_{g1} + \alpha_1 g_{e1}}{g_s} + \frac{(g_s + g_{g1} + \alpha_1 g_{e1})^2 + Y_1^2}{g_s}$$

$$\left\{ R_{eq1} + \frac{g_{g2} + g_e}{g_{m1}^2} + \frac{R_{eq2}}{g_{m1}^2} \left[\left(g_{g1} + g_{e1} + \frac{1}{r_{p2}} \right)^2 + Y_2^2 \right] \right\}$$

12 Reference 9, p. 657.

where g_g is the parallel combination of the network and ohmic losses in the tube,

g_e is the conductance electronic or transit-time loading,

g_s is the transformed source conductance,

is the ratio $\frac{Ta}{T}$ (conversion of a noise source to the standard temperature T); it is approximately 5 if the transit angle is less than one radian,

R_{eq} is the mutual conductance of the tube,

r_p is the plate resistance of the tube, and

Y is the input susceptance.

Usually $\frac{1}{r_{p2}} \gg g_{g_1} + g_{e_1}$, therefore, the shot noise contributed by the second stage is $\frac{1}{\mu_1^2}$ that of the shot noise of the first tube. The term $\frac{Y_2^2}{g_n^2}$ will be small unless the frequency is greatly different than the midband. If $\frac{g_s}{g_{m1}} \gg 1$, the contribution of the second stage thermal noise is negligible in comparison to the first tube.

The first tube and its operating conditions should be carefully chosen. The choice of the second tube depends very little on its equivalent resistance and it is usually made solely on its low cathode-plate capacity. If C_{pk} is large, neutralization is necessary; although it is usually quite uncritical. The second stage should have a fairly high transconductance because the stability of the first stage depends upon the heavy load in its plate circuit.

There are several types of input coupling networks that can be used to obtain the required bandwidth. A good discussion of these will be found in Reference 24, pp. 682-694.

Table I -- Equivalent Noise Resistances of Receiving Tubes[12]

Tube	gm, umhos	R_{eq}, ohms	C_{in}, mmf	C_{out}, mmf
Triode amplifiers				
6AC7	11,250	220	11.0	4.0
6AK5	6,670	385	4.0	2.0
6C4	2,200	1,140	1.8	1.3
6F4	5,800	430	2.0	0.6
6J4	12,000	210	2.8	0.2
6J5	2,600	960	3.4	3.6
6J6*	5,300	470	2.2	0.4
6SC7*	1,325	1,890	2.2	0.3
6SL7*	1,600	1,560	3.2	3.6
6SN7*	2,600	960	2.9	1.0
7F8*	5,650	440	2.8	1.4
9002	2,200	1,140	1.2	1.1
Sharp cutoff pentodes				
1L4	1,025	4,300	3.6	7.5
6AC7	9,000	720	11.0	5.0
6AG5	5,000	1,640	6.5	1.8
6AJ5	2,750	2,650	4.1	2.0
6AK5	5,000	1,880	4.0	2.4
6AS6	3,500	4,170	4.0	3.0
6SH7	4,900	2,850	8.5	7.0
6SJ7	1,650	5,840	6.0	7.0
9001	1,400	6,600	3.6	3.0
Remote cutoff pentodes				
1T4	750	20,000	3.5	7.3
6AB7	5,000	2,440	8.0	5.0
6SG7	4,700	4,000	8.5	7.0
6SK7	2,000	10,500	6.0	7.0
9003	1,800	13,000	3.4	3.0

* One unit of a dual triode tube.

12 Reproduced from Reference 24, p. 636.

Table II -- Comparison of Various Single-Triode Input Circuits[13]

Input tube	Advantages	Disadvantages
Grounded cathode	1. Highest available power gain, hence maximum possible reduction of 2nd-stage noise. 2. Output conductance $1/r_p$ of same order of magnitude as optimum source conductance of second stage. 3. Equivalent loss conductance equal to that of grounded plate. 4. Highest voltage gain.	1. Tendency to instability with large voltage gain, but easier to neutralize and more stable with small voltage gain than grounded plate.
Grounded plate	1. High output conductance, hence easy to get wide-band interstage coupling. 2. Induced grid noise contribution slightly less than in alternative configurations. 3. Bandwidth of input circuit greater than in grounded-cathode case because of lower input capacity.	1. Variation of grid-cathode capacity with grid bias makes neutralization difficult. 2. Tendency to instability, particularly with small G_L, even if grid-cathode capacity is resonated out. 3. Available power gain much lower than in grounded cathode case.

13 Reproduced from Reference 24, p. 652.

Table II, Contd.

Grounded grid	1. High stability due to cathode feedback and low plate-cathode capacity. 2. Large input conductance giving wide transfer band-pass characteristics for input network.	1. Low available power gain. 2. Critical dependence of first-stage noise figure on output circuit loss. 3. Greatest equivalent loss conductance. 4. Low output conductance, which combined with point 1 means second-stage noise contribution importance.

CHAPTER IV

MEASUREMENT OF NOISE FIGURE

Introduction

Noise-figure measurements give a simple and effective way of checking the ability of an amplifier to detect weak signals. The method of making the measurements is very simple and can be carried out very quickly and precisely if the proper equipment is available.

Diodes operating with temperature-limited emission may be used as a standard source of noise. Such diodes should, preferably, have either a tungsten or a thoriated tungsten filament because it is difficult to keep the emitted current from an oxide-emitting surface constant under temperature-limited conditions of emission. The noise is due to shot effect. With proper design the diode can be used as an absolute standard up to about 300 mc. Silicon crystals of the type used for rectifiers and mixers generate considerable noise when direct currents are passed through them in the "reverse" direction.

Temperature-Limited Diodes as Noise Generators

A temperature-limited diode, that is, one whose plate voltage is high enough to saturate the plate current, acts like a generator of noise current due to "shot effect". The root-mean-squared current, $\overline{i_n^2}$, is given by

$$\overline{i_n^2} = 2eIB$$

where e is the charge of the electron = 1.60×10^{-19} coulomb,

I is the direct current through the diode in amperes, and

B is the bandwidth of the device being used to observe the noise. It must be remembered that the above equation does not hold for oxide-coated cathodes because of "flicker effect".

High-Impedance Noise Generator

A simple noise generator, as shown in Figure 10, consists of a diode connected in parallel with a resistor, R_a, equal to the resistive component of the signal-source impedance and a reactance simulating the parallel reactance of the signal source. The plate voltage is made high enough to saturate the diode for all cathode temperatures used. Means are provided for varying the cathode temperature and for reading the average or d-c diode current, I_a.

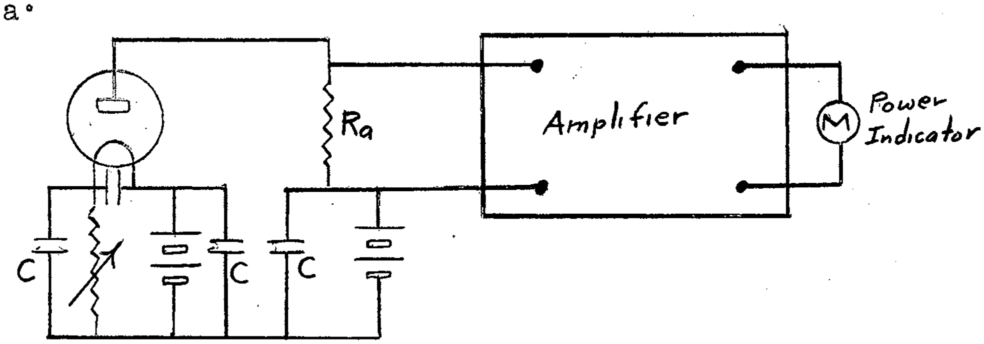

Figure 10.

The noise figure is measured by first observing the indication of the output meter with the diode filament circuit open. Under these conditions the noise output is the equivalent of that produced by the network and the source. If the meter does not match the network, it will not read the actual "available output noise power" but a quantity proportional to it. Whether the output is matched or not does not make any difference to the noise figure. This is obvious from the equation for the noise figure,

$$F = \frac{3.18 \times 10^{-19} I_a R_a}{4kT}$$

If the value 288°K is chosen for T, and if I_a is in amperes and R_a in ohms, the noise figure becomes

$$F = 20\ I_a R_a.$$

This equation is a very important one, for it gives the noise figure in terms of two easily measurable quantities, I_a and R_a.

This noise generator is very useful provided that:[1] (1) the noise generator can be connected in place of the signal source with very short leads and (2) the parallel reactance of the noise generator can be made entirely equivalent to that of the signal source over the entire range of frequencies under consideration. There is always some capacity across the diode. If it is smaller than the required capacity a padding condenser may be added. In a large percentage of the cases the desired reactance is that of a capacitor smaller than the distributed diode capcity. It is possible, but not too satisfactory, to resonate out the desired capacity by an inductance. These difficulties can be largely avoided by the use of a "low-impedance" or "matched line" noise generator.

Matched-Line Diode Noise Generator[2]

As shown in Figure 11 this generator consists of a noise diode feeding a lossless line of characteristic impedance, R_o, with a termination at the generator end and a matching network

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[1] Reference 24, p. 702.

[2] Reference 24, p. 702.

consisting of R_1, R_2, and X selected so that the internal impedance of the entire device equals that of the signal source.

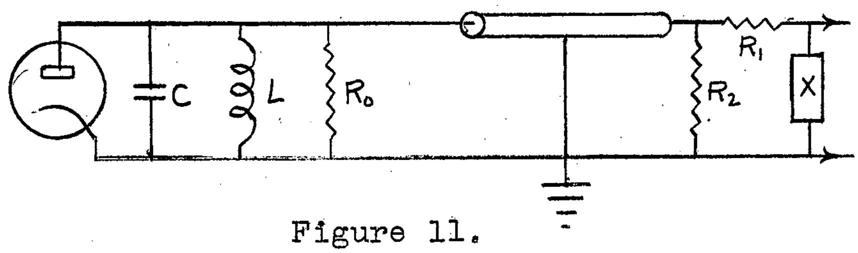

Figure 11.

C is the distributed capacity of the diode, and L resonates with C at the band-center of the amplifier to be tested. Inasmuch as R_o is a resistor of low value, the bandwidth is large compared to usual amplifier bandwidths. Since the line is terminated in its characteristic impedance at the generator end, the calibration is then independent of its length.

Measuring R_{eq} of a Tube [3]

The measurement of R_{eq} can be accomplished by associating the tube with an amplifier and power-output meter. With the grid shorted, a certain noise output will be obtained. Resistance may now be introduced into the grid circuit until the power output doubles, care being taken not to disturb bandwidth relations. If the noise in the amplifier following the tube is ignored, the resistance which doubles the output noise power is the R_{eq} for the tube operating under the conditions of the experiment. It is also assumed that no additional noise power other than thermal will be induced into the resistor because of conditions in the grid circuit of the tube.

Another method involves the substitution of a resistor for

[3] Reference 8.

the noise behavior of the tube at its plate. The equivalent resistance, R_{eq}, for the grid side of the tube may be obtained from the resistance R_{eqp}, which gives the same noise as the tube with the grid grounded, by the use of the following expression,

$$R_{eq} = \frac{R_{eqp}}{\mu^2} \frac{R_o + r_p}{R_o + R_{eqp}},$$

where R_o is the interstage or coupling resistance between the tube and the following grid circuit, which is not removed in determining R_{eqp}, and μ is the amplification factor of the tube.

<u>Construction of Noise Generators</u>

The construction of noise generators is not difficult. The generator should be completely shielded with its power leads filtered to prevent pickup of extraneous signals, especially from the output stage. Since the diodes are operated under temperature-limited conditions, the plate-supply voltage need not have good regulation.

Under temperature-limited conditions, the plate current varies very rapidly with filament voltage. If the filament is operated from the a-c power mains it is necessary to provide a constant-voltage regulator.

If batteries are used to heat the filament, then two rheostats should be used to control the filament voltage, one for rough control and the other for fine control. Sturdy construction of the filament circuit is important and, if a-c is employed on the filaments, the control switches and rheostats should be placed in the primary of the filament transformer.

Noise Power Measuring Devices

Normally output indicators present very little difficulty, and a vacuum-tube voltmeter or cathode-ray oscilloscope can be used with only ordinary precautions. The choice of an output measuring device depends upon whether or not accuracy can be sacrificed for simplicity.

Attenuator and Post Amplifier[4]

The attenuator and post amplifier combination shown in Figure 13, is the most accurate method of comparing amplifier outputs. Although it is more complex than the other schemes described here, it is very rugged and can be operated successfully by relatively untrained personnel.

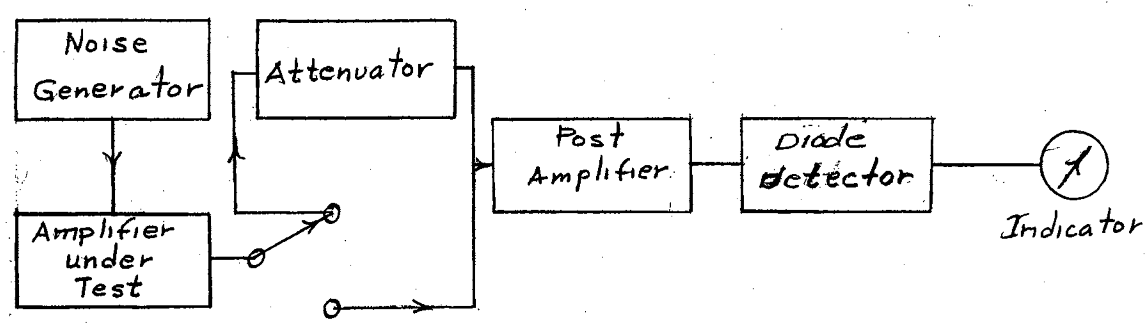

Figure 13.

The procedure consists of first to observing the output reading with the attenuator out and the noise generator off. Then the attentuator is switched on and the noise generator ad-

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4 Reference 24, p. 709.

justed to give the same reading on the output meter as before. If the attenuator has a power ratio of 2, the available power of the noise generator is that which is required to double the amplifier output power. No harm will be caused by overloading the post amplifier because it operates at constant-signal level.

The post amplifier should have the same center frequency as the amplifier under test and a bandwidth somewhat greater. It should be very well shielded and operated from its own power supply to prevent any feedback to the amplifier under test. It is convenient to gang the attenuator switch to the noise generator plate supply. However the two switches should be well isolated to prevent direct coupling from the noise generator to the post amplifier. The detector is a common diode receiver tube connected in the conventional manner.[5] A meter indicating 1 ma. full scale in series with the diode-load resistor is adequate for most measurements. However the overall sensitivity can be increased by using a meter of 30 to 100 microamperes full scale. A dry cell and a variable resistance connected in series should be placed across the meter to buck-out the "dark" current of the diode.

The attenuator and its immediate circuits, as shown in Figure 14, are designed with these objectives: (1) to make the calibration independent of the condition of impedance mismatch at the output of the amplifier and (2) to have a value of power attenuation as near 2 as possible.

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[5] Reference 26, chapter 7 and reference 24, p. 713.

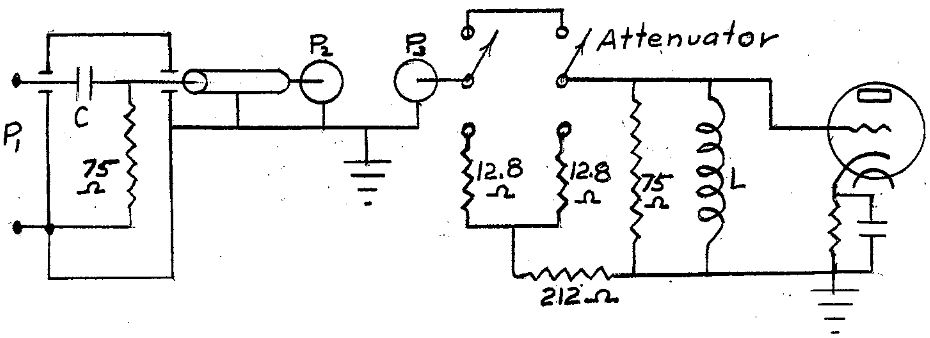

Figure 14.

If the characteristic impedance of the cable is other than 75 ohms, the values of the resistors are to be multiplied by the ratio of the characteristic impedance to 75 ohms. The 1/2-watt resistors for the attenuator are selected by a Wheatstone Bridge to be within 2 percent of the indicated values.[6]

The terminating resistor, R, and the inductance, L, which is selected to resonate with the input capacity of the tube, are designed so that the same impedance, at the center frequency of the amplifier, is measured across the pins at P_3 with both positions of the attenuator switch. If this condition is met, the calibration of the attenuator is independent of the impedance presented by the cable. The plug, P_1, fits into the tube socket of the amplifier, which is assumed to have single-tuned circuits, and is in parallel with the load of the tube previous to the tube displaced by the plug. The low impedance presented to the load by the plug causes the bandwidth of this circuit to be large. An alternative procedure is to omit the resistor in the plug, P_1,

6 If none are available, a portion of a resistor of lower value may be filed away and the filed place covered with a coat of lacquer.

by inserting between P_2 and P_3 a T- or pi-attenuator pad having a 75 ohm characteristic impedance and at least a 10 db attenuation. If the power ratio is not 2 then there should be applied a correction

$$W_n = \frac{W_1}{a-1}$$

where W_n is the true available power,

W_1 is the measured available power, and

a is the power ratio or the attenuator.

<u>Attenuator Calibration by Means of a Noise Generator</u>

An attenuator may be calibrated with a noise generator with extreme precision using only the power measuring device to indicate that the power has not changed.

First, a direct current I_1 is passed through the noise generator so that the noise-output power of the amplifier with the current I_2 flowing and the attenuator out and the noise generator off. If W_1 is the available noise generator power corresponding to I_1, then

$$W_n+W_1 = aW_n.$$

Next the noise-generator current is adjusted to a value I_2 such that with I_2 flowing and the attenuator in, the noise output power of the amplifier is the same as with the attenuator cut out and the current I_1 flowing. If W_2 is the available noise-generator power corresponding to I_2, then

$$W_n+W_2 = a(W_n+W_1).$$

Subtracting the last two equations gives

$$a = \frac{W_2-W_1}{W_1}.$$

Since W_1 and W_2 are proportional to I_1 and I_2, with the same proportionality constant, then

$$a = \frac{I_2 - I_1}{I_1} = \frac{I_2}{I_1} - 1$$

The precision with which the direct currents can be measured determines the accuracy of a.

Noise Figure Measurement with Gain-Control

A method developed by M. C. Waltz[7] allows field measurements to be made quickly and with fair accuracy. The method requires a noise generator and an output meter, usually a d-c voltmeter connected across the detector load resistor. The amplifier must also have a gain control.

First, the deflection of the meter is noted with no noise-generator current. Then an arbitrary current I_1 is passed through the generator and the output meter read a second time. Second, with the current I_1 flowing, the amplifier gain control is set so that the output meter returns to its first reading. Finally, the noise generator current is raised to a value I_2 so that the output current reads the second reading. If W_1 and W_2 correspond, respectively, to I_1 and I_2 then,

$$\frac{W_2 + W_n}{W_1 + W_n} = \frac{W_1 + W_n}{W_n}$$

Solving for W_n

$$W_n = \frac{W_1^2}{W_2 - 2W_1}$$

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7 Reference 24, p. 714.

and there results,

$$F = \frac{W_n}{kTB} = \frac{\left(\frac{W_1}{kTB}\right)^2}{\frac{W_2}{kTB} - 2\left(\frac{W_1}{kTB}\right)} .$$

For a given temperature-limited diode noise generator the ratio $\frac{W_1}{kTB}$ and $\frac{W_2}{kTB}$ have the form MI_1, and MI_2 where

$$M = 20R_a$$

for a high-impedance noise generator, and where

$$M = \frac{R_o^2}{R_a\left(1+\frac{R_o}{R_a}\right)}$$

for a matched-line noise generator at $T = 290^{\circ}K$.

Therefore the noise figure becomes,

$$F = \frac{MI_1^2}{I_2-2I_1} .$$

Some error is usually found due to the denominator being small compared to I_1 or I_2.

CHAPTER V

SUPERHETRODYNE RECEIVERS

Introduction

The superhetrodyne method of radio reception is the most widely used system. A schematic diagram for a receiver of this type is shown in Figure 15. It overcomes the difficulty of obtaining adequate carrier frequency amplification and constant selectivity over a given range.

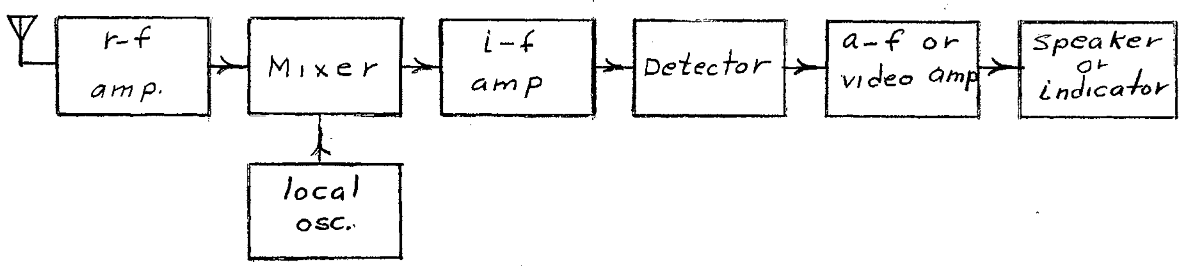

Figure 15.

For a given local-oscillator frequency, there are two possible radio frequencies that will combine to form the intermediate frequency. To suppress one of these possible r-f channels it is customary to place in front of the mixer a r-f amplifier tuned to the desired radio frequency. The r-f amplifier will also aid in eliminating most of the reradiation of the local oscillator and increases the overall signal-to-noise ratio since the mixer has a high noise figure. The local oscillator will contribute some of the noise. It should be noted that a r-f amplifier is not practical at microwave frequencies.

The mixer, by the hetrodyne process, changes the incoming signal with its modulation sidebands to a particular intermediate frequency with similar modulation sidebands. Since most of the gain may be obtained in the fixed-frequency amplifier, the selectivity and gain of the receiver are essentially independent of the radio frequency. The tuning control, with the feature of gang-tuning of local oscillator and r-f amplifier, is much simpler than that of the gang-tuned series r-f amplifiers. For reception of VHF waves, the receiver gain is obtained much easier at a relatively low intermediate frequency.

The second detector is nearly always a diode detector circuit and associated with it is the automatic gain control, agc, which controls the gain of the r-f and i-f amplifiers. The audio or video circuits which follow, amplify the intelligence contained in the carrier to a usable level.

R-F Amplifier

The r-f amplifier has been treated rather thoroughly in Chapter III and will not again be discussed here.

Mixer

The mixer, with the aid of the local oscillator converts the incoming signal to a low fixed intermediate frequency. Herold[1] gives a simple method of obtaining the conversion conductance. It should be noted that a mixer can be considered as a variable amplifier.[2]

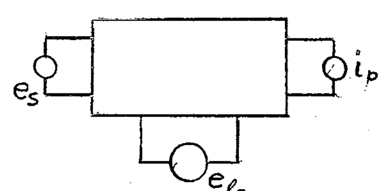

Figure 16.

[1] Reference 11.

[2] Reference 19.

As in Figure 16, a local oscillator voltage, e_{lo} is applied to one pair of terminals and a very small signal voltage, e_s, is applied to another pair. The current flowing between the other two terminals is designated as i_p. At any instant of time this current is a function of the local oscillator and signal voltages. Since the signal voltage is small the current is

$$i(t) = i_o(t) + \frac{\partial i}{\partial e_s} e_s(t)$$

where $i(t)$ is the total current at a given time,

$i_o(t)$ is the current which would flow if e_s were zero,

$e_s(t)$ is the current increment due to the signal voltage.

Since

$$\frac{\partial i}{\partial e_s} = g_m(t)$$

and since

$$e_s = E_s \cos \omega t$$

the equation may be rewritten as

$$i(t) = i_o(t) + g_m(t)E_s \cos \omega t.$$

Both terms are periodic with the local oscillator but $i_o(t)$ contains no terms that are influenced by the signal voltage, therefore it will contain no sum and difference terms.

Since only the term $g_m(t)E_s \cos \omega t$ contains the signal voltage, the other term may be dropped as far as output signal is concerned, therefore

$$i_s = g_m(t)E_s \cos \omega t.$$

Since $g_m(t)$ is periodic, it can be expressed in terms of the Fourier series

$$g_m(t) = a_o + a_1 \cos pt + a_2 \cos 2pt + a_3 \cos 3pt$$

where

$$a_o = \frac{1}{T}\int_o^T g_m(T)dt$$

$$a_n = \frac{1}{T}\int_o^T g_m(T)\cos npt\ dt \qquad n\neq o$$

p = oscillator frequency.

Due to symmetry the sine terms will be absent. After substitution then results,

$$i_s = (a_o + a_1 \cos pt + a_2 \cos 2pt + a_3 \cos 3pt + \ldots)$$
$$E_s \cos \omega t,$$

$$i_s = E_s(a_o \cos \omega t + \frac{a_1}{2} \cos (p + \omega)t + \frac{a_1}{2} \cos (p - \omega)t +$$
$$\frac{a_2}{2} \cos (2p - \omega)t + \ldots)$$

For the usual mixer the output circuit is tuned to the difference-frequency, therefore

$$i_{(i-f)} = \frac{a_1}{2} E_s \cos (p - \omega)t$$
$$= E_s \frac{\omega}{2\pi}\int_o^{\frac{2\pi}{\omega}} g_m \cos p\ tdt,$$

and hence by definition, the conversion transconductance is

$$g_{cn} = \frac{\omega}{2\pi}\int_o^{\frac{2\pi}{\omega}} g_m(t) \cos npt$$

and therefore

$$i_{(p-\omega)} = E_s g_{c1}.$$

If a harmonic of the oscillator is used, there results

$$i_{(p-n\omega)} = E_s g_{cn}.$$

The oscillator itself need not contain harmonics because the harmonics are generated in the tube due to the nonlinearity of $g_m(t)$.

The value of g_{cn} can be computed graphically by the Fourier analysis of the signal-electrode-to-output-electrode conductance vs. the oscillator-electrode voltage. By careful selection of the operating point and amplitude of the oscillator voltage the spurious responses due to the harmonics of the oscillator frequency can be kept to a minimum. This may be checked by the Fourier analysis of g_{cm} for the different harmonics of the oscillator voltage and by comparing them to the fundamental.[3]

If the cathode is not at ground potential, as in a 6SA7 with a self-excited Hartley oscillator, the signal-grid-transconductance curve must be taken with the cathode and oscillator-grid potential varied simultaneously and in their correct ratio as determined by the ratio of the cathode turns to the total turns of the coil which is used. Since the conversion transconductance is approximately proportional to the peak value of signal-grid transconductance, it is often sufficiently accurate to disregard the a-c variation of the cathode potential. This can be done by simply shifting the signal bias in the negative direction by the amount of the peak value of the alternating cathode voltage.

<u>I-F Amplifier</u>

The i-f amplifier operates at a fixed frequency. The value

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[3] Reference 2.

depends on several factors.[4] The advantages of a low intermediate frequency are,

1. for a given bandwidth the relative detuning caused by tube capacity variation is less;

2. the optimum noise figure attainable with a given tube is lower; and

3. the input loading and Miller effect are less.

The advantages of a high intermediate frequency are,

1. the image rejection is better;

2. the tuning coils and bypass condensers are smaller; and

3. a greater separation between intermediate frequencies and video frequencies is obtained thereby reducing the possibility of trouble arising from transmission of i-f components of the rectified signal into the video amplifier, and

4. the reproduction of the waveform is improved.

The types of i-f amplifiers interstage-coupling most commonly used are:

1. Synchronous single-tuned.

2. Staggered single-tuned.

3. Double tuned.

4. Inverse-feedback.

A simple resume of the different types will be given rather than a detailed analysis. An excellent treatment of the subject is given by Wallman.[5]

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[4] Reference 26, pp 158-159.

[5] Reference 24.

Each type has its advantages and disadvantages with respect to:

1. Efficiency, i.e., gain-bandwidth product.
2. Construction simplicity.
3. Noncriticalness of adjustment.
4. Ease of gain control and gain stability.
5. Selectivity.

1. Gain-bandwidth Product. The synchronous single-tuned amplifiers have gain-bandwidth products substantially smaller than any of the other amplifier types. Flat-staggered pairs have somewhat smaller gain-bandwidth products than transitionally coupled equal-Q double-tuned circuits,[6] while flat-staggered triples have slightly larger gain-bandwidth products. Transitionally coupled one-side-loaded double-tuned circuits have an advantage of exactly 2 over transitionally coupled equal-Q cir-

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[6] Transitional coupling can be expressed as

$$k_{trans} = \sqrt{\frac{1}{2}\left(\frac{1}{Q_a^2} + \frac{1}{Q_s^2}\right)}$$

for a value of k less than k_{trans}, the curve of transfer impedance against frequency shows only a single maximum; but if $Q_a \neq Q_s$, k_{trans} is greater than k_{crit}. By definition

$$Q_a = \frac{\omega_o C_a}{G_a}$$

and

$$Q_s = \frac{\omega_o C_s}{G_s},$$

and therefore

$$k_{crit} = \frac{1}{\sqrt{Q_a Q_s}} .$$

In most cases Q_s is large enough so that k_{trans} may be written as

$$k_{trans} = \frac{1}{Q_a \sqrt{2}} .$$

cuits, and in turn, surpass the one-side-loaded circuits. Plate-grid resistive feedback amplifiers with single-tuned circuit terminations, for example the usual inverse-feedback bandpass amplifiers, are exactly equivalent in gain-bandwidth product to stagger-tuned amplifiers, except that the effective transconductance is lower.

2. Simplicity. The simplest type of amplifier by far is one that employs synchronous single-tuned circuits. The tuning coils are as simple as possible and are identical from stage to stage.

Stagger-tuned and inverse-feedback amplifiers also employ very simple coils. Inverse-feedback amplifiers have the special requirement, however, that for satisfactory results the feedback resistors must have very small end-to-end capacity, about ten times less than that of the usual half-watt resistor.

Double-tuned coils are considerably more complicated than single-tuned coils. Accurate control is required of the spacer controlling mutual inductance as well as the primary and secondary inductance. This greater coil complexity is in many cases the only significant disadvantage of double-tuned circuits.

3. Tuning Stability. Synchronous single-tuned stages are extremely noncritical in tuning, while staggered pairs are less critical than staggered triples. Equal-Q double-tuned circuits are much less critical than double-tuned circuits loaded on one side only, and it has been found that stagger-damped circuits employing one-side-loaded double-tuned circuits are extremely critical. It is thought that the equal-Q double-tuned circuits

are less critical than any other except the synchronous single-tuned circuits.

4. Gain Control and Gain Stability. Except for inverse-feedback amplifiers, the individual stages of any of the amplifiers can be arbitrarily gain-controlled. Since the usual gain control varies the negative bias to the grid, the g_m of the tube is varied and, in an inverse-feedback amplifier, the bandwidth would change. The feedback amplifier has a little better gain stability.

Second Detector[7]

The usual choice of the second detector is a diode operated as a linear rectifier. The theory of this circuit can be found in any standard engineering text and will not be given here.

Practical Construction Considerations

R-f and i-f amplifiers. Great attention to detail is needed to build a stable amplifier having a gain of 100 decibels or more, and which covers a wide band of frequencies. A positive feedback factor as small as 10^{-6} anywhere in the band is very serious.

The important points to watch are:

1. Bad ground paths,
2. Waveguide feedback,
3. Inadequate bypassing of heaters, and the B+ gain control.

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7 Reference 6, chapter 7 for example.

1. Bad ground paths. At frequencies above 30 mc the chassis no longer can be considered an equal-potential ground plane of zero inductance, and the inductance of short lengths of wire begin to have appreciable effect. Good constructional practice requires that each stage should have all of its circuit components grounded in one place with short and low inductance connections. The stages should be as close together as possible to reduce stray capacity. All of the stages should be arranged in a straight line or in such a way that the ground currents of the stages are kept isolated. The inductances should be oriented so that currents induced in the chassis by their fields are a minimum. It is also necessary to locate the components so that there is no undue electrostatic coupling. The ground circuits for elements which are not part of the i-f circuit should be separated from the circuit ground. The grounding of the shells of the 6AC7's should be carried out with care since the connection between the tube pin and the shell is extra long. The heater circuit should be carefully isolated so that feedback from the input to the output terminals is a minimum. The wiring and grounding of the heater circuit should be arranged so that the wiring and chassis current do not form a large loop that intercouples with the signal currents of the amplifier. The coaxial connections to a high-gain amplifier must be made by spreading the braid in circular fashion with grounding all around the perimeter. This is important even when making measurements on the amplifier.

2. Waveguide feedback. Let us now consider the electric and

magnetic fields inside the amplifier enclosure, inasmuch as the fields and the current in the boundary are directly related by Maxwell's equations. Using the general field theory we may consider the enclosure as a waveguide and use some of the wave guide concepts and results. The cutoff frequency, f_c, in terms of the width of the guide is:

where ω is the $2\pi f$ and

$$f_c = \frac{c}{2\omega}$$

c is the velocity of light.

If the frequency used is very much smaller than the cutoff frequency the attenuation is 27.3 db over a distance equal to the width of the guide. Any box type chassis that is likely to be encountered will probably fall in this catagory. If the preceding assumption is not adequate to a particular case, a post making good electrical connections with the top and bottom of the box may be used to short out the principal mode.

3. Inadequate bypassing. It should be born in mind that a bypass condenser is actually the series combination of its lead inductance and its capacitance.

An especially effective means of bypassing involves the use of "series resonant" bypass condensers, that is, a condenser whose connecting leads resonate with the capacitances at the i-f frequency. It is important to achieve this series resonance by means of a large C and a small L, otherwise the low impedance to ground will hold only over a small part of the band. For example a 2000 mmfd. condenser with a total lead length of only 1/6 inch has a bypass impedance to ground of less than 1/2 ohm over a frequency range of 50 to 72 mc.

Parasitic oscillations may be experienced at about 500 or 600 mc. in i-f amplifiers using 6AK5's with high-Q or silver-mica button plate and screen bypass consensers. These parasitics are avoided by inserting 10-ohm carbon resistors in the paths between the B+ and the screen pins.

CHAPTER VI

CONCLUSIONS

Introduction

In the very high frequency band the average atmospheric noise is not very great and the man-made static is usually the controlling factor. Since the direction-finder is located in an isolated location, the receiver noises will be the controlling factor, but if the receiver has a very low noise factor then the noise generated in the antenna is the controlling factor. The ultimate limiting factor, therefore, is the signal-to-noise ratio in the antenna.

Antenna

So much is known and understood about antennas that most of the development work can be, and should be, done on paper. Experiment should be required only for the final engineering specifications. Before attempting adjustments, the reference material should be studied thoroughly. It is almost impossible to devise direct means for measuring the current and its phase angle at frequencies around and above 150 mc. This leaves only the actual radiation pattern to give indications of incorrect distribution and phasing of the current.

Receiver

The receiver was aligned by the use of the General Electric Model ST-4A Sweep Generator and Model ST-5A Marker Generator. These instruments were designed mainly for alignment of television sets but a sweep may be obtained in an unmarked portion of the dial and the harmonics of the marker generator may be used to set the band limits of the receiver. Care must be taken not to use the image frequency of the receiver. The bandwidth of the receiver was observed to be about 100 kcs. which may be too narrow for this type of work. The noise figure was not measured since a suitable noise standard was not available.

In order to couple the Wallman cascode amplifier into the receiver a length of 50 ohm coaxial line was used. Since the output impedance of the sweep generator was variable a 50 ohm type "T" resistive pad was connected between the receiver and the signal generator. A grid dip meter was used to obtain the proper frequency range of the local oscillator. In order to make the initial adjustments on the r-f section the grid dip meter was used to adjust the coils to the proper frequency. The sweep generator along with the marker generator was used to check the operation over the band of frequencies to be covered. No more than the usual difficulties were encountered in the alignment. The mixer did oscillate when a signal was applied but this trouble was removed by rearranging the plate lead.

The operation of the Wallman cascode amplifier was also checked with the sweep generator after the initial alignment with the grid dip oscillator. The signal was first applied to the receiver and the amount of deflection on the oscilloscope noted. Second, the Wallman circuit was placed in the circuit and the deflection on the oscilloscope compared with the first deflection. The only method of coupling that produced any gain was by obtaining series resonance of the coaxial line connection on the receiver side. This was accomplished by adding a capacitor between the loop and the center conductor of the coaxial line. The value of capacitance is very small and it was obtained by twisting two insulated wires together. The coupling was tried with and without the coaxial line and very little difference in gain was noted. The value of capacitance required only a small change.

Recommendations

There is still considerable research that should be conducted on the antenna. The elements should possibly be changed to conform with the predicted theory presented in Chapter 2. This should result in an improvement of the reception of very weak signals.

The author does not believe that the receiver has enough bandwidth to observe the waveforms and pass enough energy to give a good indication on the recording meters. Since the amount of energy received is proportional to the bandwidth, it follows that a wide bandwidth is essential. A bandwidth of 30 mc. for the i-f amplifier is suggested. The second detector should be followed by a video amplifier which feeds a peak vacuum tube voltmeter and an oscilloscope with a wide frequency range. Obviously the receiver input should be designed to have the lowest possible noise figure since, at these very high frequencies, the amplifiers are inclined to be noisy unless precautions are taken to reduce the noise. The r-f amplifier should be mounted on the same chassis as the rest of the receiver so that coupling difficulties will be minimized.

As soon as satisfactory results are obtained in the receiver and the antenna signal-to-noise ratio improved and it is suggested that there should be added a PPI indicator similar to the type used in the radar counter measure direction-finders.

Remarks

The basic fundamentals must be thoroughly understood before a satisfactory design can be formulated. It is the author's hope that this paper will present some of the basic knowledge required and will serve as a guide and incentive for those who will carry on the work.

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THESIS TITLE: Design of a VHF Direction-Finder for Tornado Detection

NAME OF AUTHOR: Howard J. Jackson

THESIS ADVISER: Herbert L. Jones

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Title of Study: Design of a VHF Direction-Finder for Tornado Detection.

Number of Pages in Study: 62

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Scope of Study: The original design of a system for the determination of the direction of arrival of atmospheric electrical disturbances involved the study of many articles and texts. Considerable time was spent on obtaining as much information as possible on antennas and low-noise amplifiers, since they are the critical parts of the direction finder. In addition to designing the direction-finder it was also necessary to construct a working model of the equipment.

Findings and Conclusions: The direction-finder employs a directional antenna to determine the direction of arrival of the electrical disturbance. The output of the antenna is fed to a Wallman cascode amplifier which preserves the original signal-to-noise by the noise generated in the succeeding stages. The low-noise amplifier is coupled to the receiver proper by a section of coaxial cable. The output indicators are a vacuum tube voltmeter which measures the automatic gain control voltage and an oscilloscope connected across the second detector load resistor to observe the waveforms and to obtain the level of weak signals.

ADVISER'S APPROVAL Herbert L. Jones